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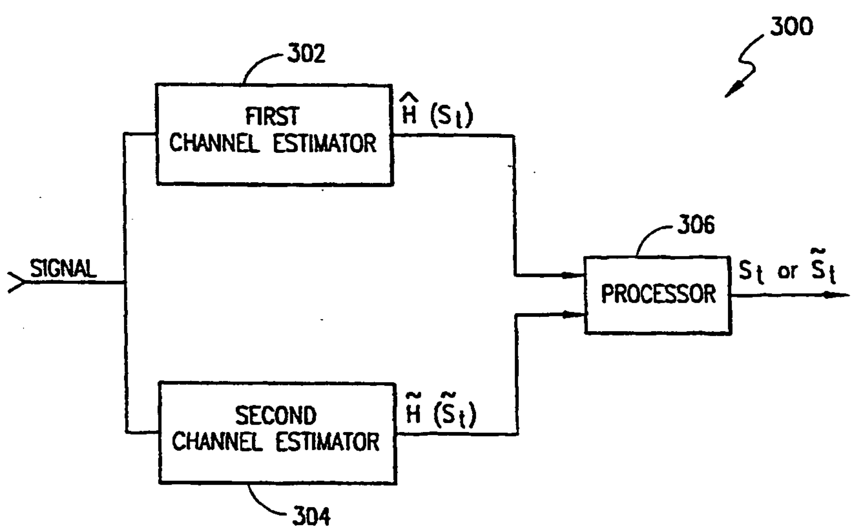
(54) Title: APPARATUS AND METHOD FOR SUBSTANTIALLY ELIMINATING A NEAR-CHANNEL INTERFERING AMPLITUDE MODULATED SIGNAL



(57) Abstract: An apparatus and method are provided that compensates for the problematic time-varying DC offset by effectively eliminating a near-channel amplitude modulated interferer from a signal. The apparatus includes a first channel estimator (302) for estimating a plurality of first channel filter taps $\hat{H}$ using a first signal model S_t, and a second channel estimator (304) for estimating a plurality of second channel filter taps H using a second signal model S_t. The apparatus also includes a processor (306) for selecting which of the first signal model S_t and the second signal model S_t is to be used or was used to substantially eliminate the near-channel amplitude modulated interferer from the received signal. The apparatus can be a mobile phone, base station, direct conversion receiver, or communications system (for example).

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APPARATUS AND METHOD FOR SUBSTANTIALLY ELIMINATING A NEAR-CHANNEL INTERFERING AMPLITUDE MODULATED SIGNAL

BACKGROUND OF THE INVENTION

Technical Field of the Invention

The present invention generally relates to the telecommunications field and, in particular, to an apparatus and method that compensates for a problematic time-varying DC offset by effectively eliminating a near-channel interfering amplitude modulated (AM) signal from a communications channel.

Description of Related Art

In the telecommunications field, one of the most significant design challenges involves the development of new direct conversion receivers that are capable of improving the demodulated quality of a signal. Traditional direct conversion receivers or homodyne receivers generally operate to demodulate an incoming signal by directly converting the incoming signal down to baseband, without the use of any intermediate frequencies, and outputting a desired signal. An example of the traditional direct conversion receiver is briefly discussed below with respect to FIGURE 1.

Referring to FIGURE 1 (PRIOR ART), there is illustrated a block diagram of a traditional direct conversion receiver 100. Basically, the traditional direct conversion receiver 100 includes an antenna 102 for receiving a signal from a transmitter 104. The received signal is filtered by a band pass filter (BPF) 106 designed to pass a desired frequency band such as the GSM (Global System for Mobile Communications) frequency band from the received signal. The filtered signal is amplified in a low noise amplifier (LNA) 108 and down-converted to a base band Inphase (I) component and a base band Quadrature (Q) component using mixers 114a and 114b, respectively, and a local oscillator (LO) 116. The local oscillator 116

outputs a frequency adapted to a carrier frequency of the received signal. The base band I and Q components are respectively filtered by first low pass filters (LPFs) 118a and 118b, converted to digital signals by analog-to-digital convertors (A/Ds) 120a and 120b, and then filtered by second low pass filters (LPFs) 122a and 122b to obtain a signal format that can be handled by a data recovery unit (DR) 124. The data recovery unit 124 operates to demodulate the received signal.

Traditional direct conversion receivers 100 have an efficient radio receiver architecture in terms of cost, size and current consumption. However, traditional direct conversion receivers 100 suffer from the well known DC offset problem that can be attributable to three different sources: (1) transistor mismatch in a signal path; (2) the local oscillator 116 outputting a signal that leaks and self-down converts to DC when passed through mixers 114a and 114b; and (3) a large near-channel amplitude modulated (AM) interfering signal leaking into the local oscillator 116 and self-downconverting to DC. Since, the resulting DC offset can be several decibels (dB) larger than the information signal, one should take care of the DC offset to be able to recover the transmitted data in the data recovery unit 124.

The DC offsets due to (1) and (2) can be assumed to be constant during one burst (i.e., a number of received symbols) and can be taken care of by adding an extra DC component to the signal model used while demodulating the transmitted data in the data recovery unit 124. This method is well known in the art. However, the DC offset due to (3) is time-varying because of the amplitude variations in the interfering signal and as such it is difficult to compensate for this particular DC offset. Two examples of how the traditional direct conversion receiver 100 can be adapted to compensate for such AM interfering signals are disclosed in WO 98/04050 and EP 0 806 841, and briefly described below with respect to FIGURE 2.

Referring to FIGURE 2 (PRIOR ART), there is illustrated a block diagram of a traditional direct conversion receiver 200 configured to compensate for AM interfering signals as described in WO 98/04050 and EP 0 806 841. The general idea disclosed in both of these documents is to add a third receiver 202 (in addition to the I and Q receivers described above) designed to compensate for the dominating AM interfering signal.

The traditional direct conversion receiver 200 excluding the third receiver 202 generally operates as the direct conversion receiver 100 described above wherein like numerals represent like parts throughout FIGURES 1 and 2. For purposes of the discussion related to the direct conversion receiver 200 of FIGURE 2, the received signal can include a wanted signal y_t and an unwanted near-channel interferer p_t. Due to nonlinear effects in the low noise amplifier 108 and the mixer 114a it can be shown that the dominated output from the second low-pass filter 122a is a wanted I component I_t and a fraction of the squared envelope of the interfering signal $a|p_t|^2$. Likewise, the dominated output from the second low-pass filter 122b is a wanted Q component Q_t and a fraction of the squared envelope of the interfering signal $b|p_t|^2$.

The third receiver 202 is designed to take into account the nonlinear effects within the low noise amplifier 108 and the mixers 114a and 114b which collectively operate to convert the interfering signal to a base band signal. The low noise amplifier 108 directs the received signal to a power detector (PD) 204 which functions to detect an envelope of the received signal. It should be noted that this detected envelope consists mainly of the envelope attributable to the near-channel AM interfering signal whenever the unwanted interferer p_t is much larger than the wanted signal y_t. The power detected signal is then converted into the digital domain by an analog-to-digital convertor (A/D) 206, filtered by a low

pass filter (LPF) 208 and fed to a control unit (CU) 210 which multiplies the detected envelopes with estimated parameters $\hat{a}$ and $\hat{b}$. The estimated interfering signals $\hat{a}|p_t|^2$ and $\hat{b}|p_t|^2$ of the distortion are respectively input to subtractors 212a and 212b and subtracted from the I and Q components to obtain "relatively clean" I and Q components. The "relatively clean" I and Q components are then input to the data recovery unit 124.

Even if the solution to the DC offset problem described in WO 98/04050 and EP 0 806 841 appears to be promising it still has disadvantages, in terms of cost and current, due to the need to implement a third receiver. Therefore, there is a need for an apparatus and method that can suppress the near-channel AM interferer in a cost and current efficient manner.

BRIEF DESCRIPTION OF THE INVENTION

The present invention is an apparatus and method that compensates for the problematic time-varying DC offset by substantially eliminating a near-channel amplitude modulated interferer from a signal. The apparatus includes a first channel estimator for estimating a plurality of first channel filter taps $\hat{H}$ using a first signal model S_t, and a second channel estimator for estimating a plurality of second channel filter taps $\tilde{H}$ using a second signal model $\tilde{S}_t$. The apparatus also includes a processor for selecting which of the first signal model S_t and the second signal model $\tilde{S}_t$ is to be used or was used to substantially eliminate the near-channel amplitude modulated interferer from the received signal. The apparatus and method can be implemented in, for example, a mobile phone, base station, direct conversion receiver, or communications system.

BRIEF DESCRIPTION OF THE DRAWINGS

A more complete understanding of the method and apparatus of the present invention may be had by reference

to the following detailed description when taken in conjunction with the accompanying drawings wherein:

FIGURE 1 (PRIOR ART) is a block diagram illustrating the basic components of a traditional direct conversion receiver;

FIGURE 2 (PRIOR ART) is a block diagram illustrating a traditional direct conversion receiver adapted to include a third receiver for DC offset compensation;

FIGURE 3 is a block diagram illustrating the basic components associated with an exemplary apparatus of the present invention;

FIGURE 4 is a block diagram illustrating in greater detail a first embodiment of the exemplary apparatus shown in FIGURE 3;

FIGURE 5 is a block diagram illustrating in greater detail a second embodiment of the exemplary apparatus shown in FIGURE 3;

FIGURE 6 is a block diagram illustrating in greater detail a third embodiment of the exemplary apparatus shown in FIGURE 3;

FIGURE 7 is a block diagram illustrating in greater detail a fourth embodiment of the exemplary apparatus shown in FIGURE 3; and

FIGURE 8 is a flowchart illustrating the basic steps of the preferred method in accordance with the present invention.

DETAILED DESCRIPTION OF THE DRAWINGS

Referring to FIGURES 3-8, there are disclosed exemplary embodiments of an apparatus 300 and preferred method 800 each of which is capable of effectively eliminating near-channel amplitude modulated (AM) interfering signals from signals in accordance with the present invention.

Although the apparatus 300 is described with respect to a direct conversion receiver (e.g., homodyne receiver) used in a digital communications system, it should be

understood that the present invention can be used within any communications device, but is especially suited for use with a mobile phone and base station. Accordingly, the different embodiments of the apparatus 300 and preferred method 800 should not be construed in a limited manner.

Referring to FIGURE 3, there is a block diagram illustrating the basic components associated with the exemplary apparatus 300 of the present invention. Basically, the apparatus 300 (described as a direct conversion receiver) enables the suppression of AM interfering signals using only the I and Q baseband signals in a predetermined manner such that no extra receiver (see the third receiver 202 of FIGURE 2) is needed, implying a cost and current efficient receiver having low complexity and high performance.

More specifically, the direct conversion receiver 300 of the present invention includes a first channel estimator 302 operable to estimate a plurality of first channel filter taps $\hat{H}$ using a first signal model S_t, and a second channel estimator 304 operable to estimate a plurality of second channel filter taps $\tilde{H}$ using a second signal model S_r. The direct conversion receiver 300 also includes a processor 306 operable to select either the first signal model S_t or the second signal model S_r that is to be used or was used to substantially eliminate the near-channel AM interfering signal from the received signal. A detailed description of how each embodiment of the direct conversion receiver 300 operates to substantially eliminate the AM interfering signal is provided below with respect to FIGURES 4-7.

Referring to FIGURE 4, there is a block diagram illustrating in greater detail a first embodiment of an exemplary direct conversion receiver 400 in accordance with the present invention. Certain details associated with direct conversion receivers are known in the industry and as such need not be described herein. Therefore, for clarity, the descriptions provided below in relation to the

direct conversion receivers of the present invention omit some elements known to those skilled in the art that are not necessary to understand the invention.

The direct conversion receiver 400 includes an antenna 402 for receiving a signal from a transmitter 404. The received signal is filtered by a band pass filter (BPF) 406 designed to pass a desired frequency band such as the GSM (Global System for Mobile Communications) frequency band from the received signal. The received signal output from the band pass filter 406 can be represented as:

$$w_t = y_t - \tilde{p}_t \quad (1)$$

$$= r_t \cos(\omega_0 t + \varphi_t) + p_t \cos(\omega_1 t + \theta_t) \quad (2)$$

where w_t is the received signal, y_t is the desired signal at carrier frequency ω_0, $\tilde{p}_t$ is the near-channel AM interfering signal at carrier frequency ω_1, and ω_0 and ω_1 are within the pass band of the band pass filter 406.

The filtered signal is amplified in a low noise amplifier (LNA) 408 and down-converted to a base band Inphase (I) component and a base band Quadrature (Q) component using mixers 414a and 414b, respectively, and a local oscillator (LO) 416. The local oscillator 416 outputs two oscillation signals LO_I and LO_Q adapted to a carrier frequency of the received signal, the two oscillation signals can be represented as follows:

$$LO_I(t) = \cos(\omega_0 t) \quad (3)$$

$$LO_Q(t) = \sin(\omega_0 t) \quad (4)$$

where LO_I and LO_Q are the oscillating signals associated with the I and Q components, respectively. The oscillating signals LO_I and LO_Q and the received signal are multiplied in the mixers 414a and 414b.

Due to the nonlinearities of the local oscillator 416 and interferer leakage (represented with the scale factors α', β'), the low pass filtered signal output from the I

component mixer 414a and a first low pass filter (LPF) 418a can be represented as follows:

$$\tilde{I}_t = LPF\left\{ \left(r_t\cos(\omega_0 t+\varphi_t)+p_t\cos(\omega_1 t+\theta_t)\right)+\alpha'_I\cos(\omega_0 t+\gamma)\right) * \left(\cos(\omega_0 t) + \beta'_I\, p_t\cos(\omega_1 t+\theta_t+\delta)\right)\right\} \quad (5)$$

$$= r_t\cos\varphi_t + \beta_I\, p_t^2 + \alpha_I \quad (6)$$

Likewise, the low pass filtered signal output from the Q component mixer 414b and a first low-pass filter (LPF) 418b can be represented as follows:

$$\tilde{Q}_t = r_t\sin\varphi_t + \beta_Q\, p_t^2 + \alpha_Q \quad (7)$$

Thereafter, the $\tilde{I}$ and $\tilde{Q}$ components are respectively converted into digital domain by analog-to-digital convertors (A/Ds) 420a and 420b and respectively filtered by second low pass filters (LPFs) 422a and 422b. And, after certain normalizations the base band components can be represented as:

$$\tilde{I}_t = I_t + a|p_t|^2 + I_{DC} \quad (8)$$
$$\tilde{Q}_t = Q_t + b|p_t|^2 + Q_{DC} \quad (9)$$

where I_t, Q_t are the wanted I and Q components, and I_{DC}, Q_{DC} are the DC components on the I and Q components, respectively. $|p_t|^2$ is the low pass filtered and sampled squared envelope of the interfering AM signal. In case of digital transmission over radio channels with intersymbol interference, such as for instance in GSM or a D-AMPS cellular systems, the wanted I and Q components can be written in a complex notation as follows:

$$I_t + jQ_t = H^T U_t + e_t \quad (10)$$

where $H = [h_0, \ldots, h_L]^T$ is a vector of complex valued channel filter taps, $U_t = [u_t, \ldots, u_{t-L}]^T$ is a vector of complex transmitted symbols, and e_t is some kind of complex valued

noise. Therefore, the complex valued base band signal for a first signal model S_t can be represented as:

$$S_t = \tilde{I}_t + j\tilde{Q}_t \quad (11)$$

$$= H^T U_t + (a+jb)|p_t|^2 + DC + e_t \quad (12)$$

$$= \sum_{k=o}^{L} \left\{h_k^I + j h_u^Q\right\} \left\{U_t^I + j U_t^Q\right\} + e_t^I + j e_t^Q \quad (13)$$

where $j = \sqrt{-1}$.

The $\tilde{I}_t$ and $\tilde{Q}_t$ components are input to a first channel estimator 424a that correlates, using the first signal model S_t, a known training sequence (TS) with the received signal S_t (which contains the same known training sequence) to determine a synchronization position and an estimate of the first set of channel filter taps $\hat{H}$. The use of the first signal model S_t in estimating the channel filter taps $\hat{H}$ is well known in the art. In fact, the first signal model S_t was the only signal model used to estimate channel filter taps $\hat{H}$ in traditional direct conversion receivers. The estimated channel filter taps $\hat{H}$ are input to a processor 426 that is described in detail below.

Generally, the present invention includes a second channel estimator 424b that uses a second signal model $\tilde{S}_t$ to estimate a second set of channel filter taps $\tilde{H}$ that are input to the processor 426 which selects the signal model S_t or $\tilde{S}_t$ that is to be used to further process the received signal. In other words, the processor 426 selects the appropriate signal model S_t or $\tilde{S}_t$ using the estimated channel filter taps $\hat{H}$ and $\tilde{H}$ and some other parameters (e.g., residuals) discussed below. Thereafter, an equalizer 428, coupled to the processor 426, uses either the first or second set of channel filter taps $\hat{H}$ or $\tilde{H}$ corresponding to the selected signal model S_t or $\tilde{S}_t$ to equalize the received signal. It should be noted that the equalizer 428 also receives $\tilde{I}_t$ from the second low pass filter 422a, $\tilde{Q}_t$ from the second low pass filter 422b, and $\tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$

from a subtractor 434 before equalizing the received signal.

More specifically, the second signal model $\tilde{S}_t$, represented as the component $\tilde{Q}_t-\frac{\hat{b}}{\hat{a}}\tilde{I}_t$ where $\hat{a}$ and $\hat{b}$ are respective estimates of a and b from equations 8 and 9, is input to the second channel estimator 424b. The second channel estimator 424b operates to correlate the known training sequence (TS) with the $\tilde{Q}_t-\frac{\hat{b}}{\hat{a}}\tilde{I}_t$ component to determine a synchronization position and an estimate of the second set of channel filter taps $\tilde{H}$. The $\frac{\hat{b}}{\hat{a}}\tilde{I}_t$ portion of the second signal model $\tilde{S}_t$ is generated using a multiplicator 432 which receives $\tilde{I}_t$ from the second low pass filter 422a, and receives the estimated parameters $\hat{a}$ and $\hat{b}$ from the processor 426. The multiplicator 432 outputs $\frac{\hat{b}}{\hat{a}}\tilde{I}_t$ to the subtractor 434 which receives $\tilde{Q}_t$ from the second low pass filter 422b and outputs the component $\tilde{Q}_t-\frac{\hat{b}}{\hat{a}}\tilde{I}_t$.

The second signal model $\tilde{S}_t$ can be represented as follows:

$$\tilde{S}_t(\hat{b}/\hat{a})=\tilde{Q}_t-\frac{\hat{b}}{\hat{a}}\tilde{I}_t \tag{14}$$

$$= \tilde{H}^T\tilde{U}_t + imag(e_t) - \frac{\hat{b}}{\hat{a}}\, real(e_t) + R_{DC} \tag{15}$$

where $\tilde{U}_t$ = [real(U_t) imag(U_t)] does not contain any distortion $|p_t|^2$, implying the elimination of the AM interferer.

The derivation to obtain equation (15) from equation (14) follows:

$$\tilde{S}_t(\frac{\hat{b}}{\hat{a}})=\tilde{Q}_t-\frac{\hat{b}}{\hat{a}}\ \tilde{I}_t = Q_t-\frac{\hat{b}}{\hat{a}}I_t + \left[b - \frac{\hat{b}a}{\hat{a}}\right]|p_t|^2+\left\{Q_{DC}-\frac{\hat{b}}{\hat{a}}\ I_{DC}\right| \quad (16)$$

where $Q_{DC}-\frac{\hat{b}}{\hat{a}}I_{DC} = R_{DC}$ (see equation 15), and assuming $\hat{a}$ = a and $\hat{b}$ = b then $|p_t|^2$ vanishes.

Equation 10 can be written as:

$$I_t + jQ_t = \sum_k \left\{h_k^I u_{t-k}^I - h_k^Q u_{t-k}^Q\right\} + j\left\{h_k^I u_{t-k}^Q + h_k^Q u_{t-k}^I\right\} + e_t^I + je_t^Q \quad (17)$$

where e_t is expressed in real and imaginary parts as $e_t = e_t^I + je_t^Q$, h^I and h^Q respectively represent the real and imaginary parts of the channel filter taps h, and u^I and u^Q respectively represent the real and imaginary parts of the transmitted symbols u_t.

Rearranging equation 17 yields

$$I_t + jQ_t = e_t^I + \left\{\sum_k h_k^I u_{t-k}^I - h_k^Q u_{t-k}^Q\right\} + j\left(\left\{\sum_k h_k^I u_{t-k}^Q + h_k^Q u_{t-k}^I\right\} + e_t^Q\right) \quad (18)$$

where $e_t^I + \left\{\sum_k h_k^I u_{t-k}^I - h_k^Q u_{t-k}^Q\right\} = I_t$ and $\left(\left\{\sum_k h_k^I u_{t-k}^Q + h_k^Q u_{t-k}^I\right\} + e_t^Q\right) = \tilde{Q}_t$

Then, substituting for I_t and Q_t in equation 16, and assuming $\hat{a} = a$ and $\hat{b} = b$,

$$\tilde{s}_t\left(\frac{\hat{b}}{\hat{a}}\right) = \sum_k \left(h_k^Q - \frac{\hat{b}}{\hat{a}} h_k^I\right) u_{t-k}^I + \sum_k \left(h_k^I - \frac{\hat{b}}{\hat{a}} h_k^Q\right) u_{t-k}^Q + imag\{e_t\} - \frac{\hat{b}}{\hat{a}} re\{e_t\} + R_{DC} \quad (19)$$

where $imag\{e_t\} = e_t^I$ and $re(e_t) = e_t^Q$.
And, in matrix form,

$$\tilde{s}_t = \left\{\left[h_0^Q - \frac{\hat{b}}{\hat{a}} h_0^I\right] \ldots \left[h_L^Q - \frac{\hat{b}}{\hat{a}} h_L^I\right], \left[h_0^I - \frac{\hat{b}}{\hat{a}} h_0^Q\right] \ldots \left[h_L^I - \frac{\hat{b}}{\hat{a}} h_L^Q\right]\right\} \begin{Bmatrix} u_t^I \\ \vdots \\ u_{t-L}^I \\ u_t^Q \\ \vdots \\ u_{t-L}^Q \end{Bmatrix} + imag\{e_t\} - \frac{\hat{b}}{\hat{a}} re\{e_t\} + R_{DC} \quad (20)$$

Equation 20 can now be written in the form of equation 15:

$$= \tilde{H}^T \tilde{U}_t + imag\{e_t\} - \frac{\hat{b}}{\hat{a}} real\{e_t\} + R_{DC} \quad (21)$$

However, since a and b (see equations 8 and 9) are not known, they have to be estimated within the channel estimator 424b using the second signal model $\tilde{S}_t$ and the

resulting DC component, R_{DC}. One way of estimating $\hat{a}$ and $\hat{b}$ for each received burst is described in the exemplary optimizing b/a algorithm which follows:

1. Set $i = o$.

2. Let $\dfrac{b}{\hat{a}} = \dfrac{b_i}{a_i}$. (22)

The start value b_0/a_0 can be based on some a priori information about these parameters, for instance some nominal nonlinear performance for the particular low noise amplifier 408 and mixers 414a and 414b. For instance, the b/a ratio can be between 1/10 to 10 then one can make a grid of N values (corresponding to i=0...N-1) between 1/10 to 10 and compute Q (b_i/a_i) for all of these values (see Equation 24 and Steps 6-7 below).

3. Find $\tilde{S}_t(\dfrac{\hat{b}}{\hat{a}}) = \tilde{Q}_t - \dfrac{\hat{b}}{\hat{a}}\tilde{I}_t$ (23)

4. Synchronize to find the best synchronization position or known symbol pattern in the received burst. For example, this can be done by correlating between the received burst and the training sequence.

5. Channel estimate to find the estimated channel filter taps $\tilde{H}$ and R_{DC} for the signal $\tilde{S}_t(\dfrac{\hat{b}}{\hat{a}})$. For example, this can be done using conventional least-squares techniques as indicated by equation 24 below:

$$Q^{\min}\left[\frac{\hat{b}}{\hat{a}}\right] = \min_{\tilde{h}_k^I, R_{DC}, \tilde{h}_k^Q} \sum_{L=1}^{N}\left[\tilde{S}_t\left(\frac{\hat{b}}{\hat{a}}\right) - \sum_{k=1}^{L}\tilde{h}_k^I u^I{}_{t-k} - \sum_{k=1}^{L}\tilde{h}_k^Q u^Q{}_{t-k} - R_{DC}\right]^2 \quad (24)$$

$$\text{store} \quad Q^{\min}\left[\frac{\hat{b}}{\hat{a}}\right] = f(i) \quad \text{and} \quad \tilde{H}(i) = \tilde{H}_{opt}\left\{\frac{b_i}{a_i}\right\}$$

where $\tilde{H}_{opt}$ is the vector that yields $f(i)$ in equation 24.

6. Set $i = i + 1$

7. Perform steps 2 - 6 repetitively until all $\frac{a_i}{b_i}$ (e.g., i=0...N-1) are used.

8. Find the lowest value, f_{min}, of all of the f(i) values. Select the corresponding $\frac{a_i}{b_i}$ value to be $\frac{\hat{a}}{\hat{b}}$, and select the corresponding $\tilde{H}(i)$ vector to be the estimated channel tap vector $\tilde{H}$.

Another way of estimating $\hat{a}$ and $\hat{b}$ for each received burst can be accomplished using various numerical search methods such as (for example):

$$\frac{b_i}{a_i} = \frac{b_{i-1}}{a_{i-1}} + f\left[Q\left(\frac{b_{(i-1)}}{a_{(i-1)}}\right)\right]$$

where the function $f\left[Q\left(\frac{b_{(i-1)}}{a_{(i-1)}}\right)\right]$ depends on the numerical search method utilized to find the optimal a and b values. For instance, the well known gradient method can be used where f is basically the derivative of Q (calculated as in Equation 24).

It should be understood that variations in the parameters a and b depends on quantities such as temperature and aging, implying a time constant which is much slower than a time constant of the radio channel. Therefore, the algorithm for finding optimal values of $\hat{a}$ and $\hat{b}$ need not be executed for every received burst, but just within some certain time intervals such as, for instance, every n:th received burst, or every k:th second.

How well the AM interfer is eliminated depends on how large $|p_t|^2$ is compared to the noise e_t and also the relation between a and b. As such, for each particular received burst there is a determination as to whether the standard

first signal model S_t or the second signal model $\tilde{S}_t$ gives a higher (e.g., highest) signal-to-noise ratio. The processor 426 selects the first signal model S_t or the second signal model $\tilde{S}_t$ based on the estimated channels filter taps $\tilde{H}$ and $\hat{H}$ and some quality parameters such as the residuals obtained in the channel estimators 424a and 430b.

Following is an exemplary way of how the processor 426 can decide whether to use the first signal model S_t or the second signal model $\tilde{S}_t$. First, compare f_{min} (see equation 24) to g_{min} where:

$$g_{\min} = \min_{\hat{h}_k} \sum_{L=1}^{N} \left(S_t - \sum_{k=1}^{L} \hat{h}_k u_{t-k} \right)^2 . \tag{25}$$

If $f_{min} < \alpha g_{min}$ (where α is an application specific design parameter) then select the second signal model $\tilde{S}_t$ otherwise select the first signal model S_t. Thereafter, the estimated channel taps $\hat{H}$ or $\tilde{H}$ corresponding to the selected signal model S_t or $\tilde{S}_t$ are input to the equalizer 428 that decodes the received signal.

A typical value of α is 1 which indicates that one selects the signal model $\tilde{S}_t$ or S_t having the higher signal-to-noise ratio. However, empirical results indicate a better performance when an α smaller than 1 (e.g., 0.2-0.95) is utilized, in which case the second signal model $\tilde{S}_t$ must have a significantly better signal-to-noise ratio than the first signal model S_t before it is chosen.

It should also be understood that the direct conversion receiver (any embodiment) is capable of operating using only the second signal model $\tilde{S}_t$, instead of having the processor 426 select which of the signal models S_t or $\tilde{S}_t$ best fits the received signal.

Referring to FIGURE 5, there is a block diagram illustrating in greater detail a second embodiment of an exemplary direct conversion receiver 500 in accordance with

the present invention. The direct conversion receiver 500 is similar to the first embodiment except that instead of using the received signal to estimate parameters $\hat{a}$ and $\hat{b}$, the direct conversion receiver 500 uses internally generated test signals c_t and d_t to estimate parameters $\hat{a}$ and $\hat{b}$.

To avoid repetition, only the components used to internally generate the test signals c_t and d_t in the direct conversion receiver 500 are described, because the direct conversion receivers of the first and second embodiments otherwise have basically the same architecture and functionality.

The direct conversion receiver 500 includes a Digital Signal Processor (DSP) 502 or an Application-Specific Integrated Circuit (ASIC) operable to digitally generate the waveforms of the base band test signals c_t and d_t. The test signals c_t and d_t include the received (desired) signal and the AM interfering signal, and can be represented as follows:

$$c_t = r_t\cos(\Phi_t) + p_t\cos(2\pi\Delta ft+\theta_t) \tag{26}$$

$$d_t = r_t\sin(\Phi_t) + p_t\sin(2\pi\Delta ft+\theta_t) \tag{27}$$

where $r_t\cos(\Phi_t)$ *and* $r_t\sin(\Phi_t)$ are the I and Q components of the desired signal, and $p_t\cos(\Delta\omega t+\theta_t)$ *and* $p_t\sin(\Delta\omega t+\theta_t)$ are the I and Q components of the AM interfering signal Δf hertz from the desired signal.

The internally generated test signals c_t and d_t are respectively filtered in low pass filters (LPFs) 504a and 504b, and input to mixers 506a and 506b that convert the base band signals c_t and d_t up to the carrier frequency using the oscillating signals LO_I and LO_Q from the local oscillator 416. The mixers 506a and 506b output their respective internally generated test signals to an adder 508 which outputs an internally generated test signal to the band pass filter 406 through a switch 510.

Thereafter, the direct conversion receiver 500 operates to estimate the parameters $\hat{a}$ and $\hat{b}$ in the same manner as described above with respect to the first embodiment, except that the internally generated test signals c_t and d_t are used instead of the signal received at the antenna 402. After estimating the parameters $\hat{a}$ and $\hat{b}$, the switch 510 is positioned to connect the antenna 402 and the band pass filter 406 to enable the further operation of the direct conversion receiver 500 (see discussion with respect to FIGURE 4).

An advantage of using the internally generated test signals c_t and d_t in this self test option is that one can design and control the interfering signal in such a way that the identification process of the parameters a and b can be easily optimized.

Referring to FIGURE 6, there is a block diagram illustrating in greater detail a third embodiment of an exemplary direct conversion receiver 600 in accordance with the present invention. The direct conversion receiver 600 is similar to the first embodiment except that instead of using the received signal to estimate parameters $\hat{a}$ and $\hat{b}$, the direct conversion receiver 600 uses a single internally generated test signal g_t to estimate parameters $\hat{a}$ and $\hat{b}$.

To avoid repetition, only the components used to internally generate the test signal g_t in the direct conversion receiver 600 are described, because the direct conversion receivers of the first and third embodiments otherwise have basically the same architecture and functionality.

The direct conversion receiver 600 includes a Digital Signal Processor (DSP) 602 or an Application-Specific Integrated Circuit (ASIC) operable to digitally generate the waveform of the base band test signal g_t. The test signal g_t includes the received (desired) signal and the AM interfering signal, and can be represented as follows:

$$g_t = r_t \cos(\Phi_t) + p_t \cos(2\pi \Delta f t + \theta_t) \qquad (28)$$

where $r_t\cos(\Phi_t)$ is the I component of the desired signal, and $p_t\cos(\Delta\omega t+\theta_t)$ is the I component of the AM interfering signal Δf hertz from the desired signal.

Thereafter, the internally generated test signal g_t is filtered by a low pass filter (LPF) 604, and input to a mixer 606 that converts the base band signal g_t up to the carrier frequency using the oscillating signal LO_I from the local oscillator 416. The mixer 606 creates two interfering signals an equal distance ($\pm\Delta f$) from the desired carrier such that the use of the desired signal $r_t\cos(\Phi_t)$ may not be needed. Therefore, the internally generated test g_t can be represented as follows:

$$g_t = p_t\cos(2\pi\Delta f t+\theta_t) \qquad (29)$$

The mixer 606 outputs the internally generated test signal to the band pass filter 406 through a switch 610. Thereafter, the direct conversion receiver 600 operates to estimate the parameters $\hat{a}$ and $\hat{b}$ in the same manner as described above with respect to the first embodiment, except that the internally generated test signal g_t is used instead of the signal received at the antenna 402. After estimating the parameters $\hat{a}$ and $\hat{b}$, the switch 610 is positioned to connect the antenna 402 and the band pass filter 406 to enable the further operation of the direct conversion receiver 500 (see discussion with respect to FIGURE 4).

An advantage of using the internally generated test signal g_t in this self test option is that one can design and control the interfering signal in such a way that the identification process of the parameters a and b can be easily optimized.

Referring to FIGURE 7, there is a block diagram illustrating in greater detail a fourth embodiment of an exemplary direct conversion receiver 700 in accordance with the present invention. The direct conversion receiver 700 is similar to the first embodiment except that instead of

selecting the first or second signal model S_t or $\tilde{S}_t$ before the equalizer 428 (see FIGURE 4) the selection of the signal model S_t or $\tilde{S}_t$ is made after first and second equalizers 728a and 728b.

The direct conversion receiver 700 includes an antenna 702 for receiving a signal from a transmitter 704. The received signal is filtered by a band pass filter (BPF) 706 designed to pass a desired frequency band such as the GSM (Global System for Mobile Communications) frequency band from the received signal. The received signal output from the band pass filter 706 can be represented as:

$$\begin{aligned} w_t &= y_t + \tilde{p}_t & (30) \\ &= r_t \cos(\omega_0 t + \varphi_t) + p_t \cos(\omega_1 t + \theta_t) & (31) \end{aligned}$$

where w_t is the received signal, y_t is the desired signal at carrier frequency ω_0, $\tilde{p}_t$ is the near-channel AM interfering signal at carrier frequency ω_1, and ω_0 and ω_1 are within the pass band of the band pass filter 706.

The filtered signal is amplified in a low noise amplifier (LNA) 708 and down-converted to a base band Inphase (I) component and a base band Quadrature (Q) component using mixers 714a and 714b, respectively, and a local oscillator (LO) 716. The local oscillator 716 outputs two oscillation signals LO_I and LO_Q adapted to a carrier frequency of the received signal, the two oscillation signals can be represented as follows:

$$LO_I(t) = \cos(\omega_0 t) \qquad (32)$$
$$LO_Q(t) = \sin(\omega_0 t) \qquad (33)$$

where LO_I and LO_Q are the oscillating signals associated with the I and Q components, respectively. The oscillating signals LO_I and LO_Q and the received signal are multiplied in the mixers 714a and 714b.

Due to the nonlinearities of the local oscillator 716 and interferer leakage (represented with the scale factors

α', β'), the low pass filtered signal output from the I component mixer 714a and a first low pass filter 718a can be represented as follows:

$$\tilde{I}_t = LPF\left\{ \begin{array}{l}(r_t\cos(\omega_0 t+\varphi_t)+p_t\cos(\omega_1 t+\theta_t)+\alpha'_I \cos(\omega_0 t+\gamma)) \\ *(\cos(\omega_0 t) + \beta'_I\, p_t\cos(\omega_1 t+\theta_t+\delta))\end{array}\right\} \tag{34}$$

$$= r_t \cos\varphi_t + \beta_I\, p_t^2 + \alpha_I \tag{35}$$

Likewise, the low pass filtered signal output from the Q component mixer 714b and a first low pass filter (LPF) 718b can be represented as follows:

$$\tilde{Q}_t = r_t \sin\varphi_t + \beta_Q\, p_t^2 + \alpha_Q \tag{36}$$

Thereafter, the I and Q components are respectively converted into digital domain by analog-to-digital convertors (A/Ds) 720a and 720b and respectively filtered by second low pass filters (LPFs) 722a and 722b. And, after certain normalizations the base band components can be represented as:

$$\tilde{I}_t = I_t + a|p_t|^2 + I_{DC} \tag{37}$$

$$\tilde{Q}_t = Q_t + b|p_t|^2 + Q_{DC} \tag{38}$$

where I_t, Q_t are the wanted I and Q components, and I_{DC}, Q_{DC} are the DC components on the I and Q components, respectively. $|p_t|^2$ is the low pass filtered and sampled squared envelope of the interfering AM signal. In case of digital transmission over radio channels with intersymbol interference, such as for instance in GSM or a D-AMPS cellular systems, the wanted I and Q components can be written in a complex notation as follows:

$$I_t + jQ_t = H^T U_t + e_t \tag{39}$$

where $H = [h_0, \ldots, h_L]^T$ is a vector of complex valued channel filter taps, $U_t=[u_t, \ldots, u_{t-L}]^T$ is a vector of complex

transmitted symbols, and e_t is some kind of complex valued noise. Therefore, the complex valued base band signal or a first signal model S_t can be represented as:

$$S_t = \tilde{I}_t + j\tilde{Q}_t \qquad (40)$$

$$= H^T U_t + (a+jb)|p_t|^2 + DC + e_t \qquad (41)$$

$$= \sum_{k=0}^{L} \left\{h_k^I + jh_u^Q\right\}\left\{U_t^I + jU_t^Q\right\} + e_t^I + je_t^Q \qquad (42)$$

where $j = \sqrt{-1}$.

The $\tilde{I}_t$ and Q_t components are input to a first channel estimator 724a that correlates, using the first signal model S_t, a known training sequence (TS) with the received signal S_t containing the same known training sequence to determine a synchronization position and an estimate of the first set of channel filter taps $\hat{H}$. The use of the first signal model S_t in estimating the channel filter taps $\hat{H}$ is well known in the art. The estimated channel filter taps $\hat{H}$ are then input to a first equalizer 728a that decodes the signal and outputs the decided signal $\hat{u}_t$ and metrics to a processor 726.

Generally, the present invention includes a second channel estimator 724b that utilizes a second signal model $\tilde{S}_t$ to estimate a second set of channel filter taps $\tilde{H}$. The second set of channel filter taps $\tilde{H}$ are input to the second equalizer 728b that decodes the signal and outputs the decided signal $\hat{u}_t$ and metrics to the processor 726. To obtain the second signal model $\tilde{S}_t$ represented as the component $\tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$, the second low pass filter 722a outputs the $\tilde{I}_t$ component to a multiplicator 730 that also receives estimated parameters $\hat{b}/\hat{a}$ from channel estimator 724b. The estimated parameters $\hat{a}$ and $\hat{b}$ are determined in the channel estimator 724b in a similar manner as described-above with respect to the channel estimator 424b of the first embodiment (see FIGURE 4 and related description). The multiplicator 730 outputs $-\frac{\hat{b}}{\hat{a}}\tilde{I}_t$ to a subtractor 732 that

also receives $\tilde{Q}_t$ and, in turn, outputs $\tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$ to the second equalizer 728b and the second channel estimator 724b.

More specifically, the second channel estimator 724b correlates, using the second signal model $\tilde{S}_t$, the known training sequence (TS) with the $\tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$ component to determine a synchronization position and an estimate of the second set of channel filter taps $\tilde{H}$. The estimated channel filter taps $\tilde{H}$ are input to the second equalizer 728b that decodes the signal and outputs the decided signal $\tilde{u}_t$ and metrics to the processor 726. Thereafter, the processor 726 selects either the first or second signal models S_t or S_r based on the metrics and decided signals $\hat{u}_t$ and $\tilde{u}_t$.

The selection between the signal models $\tilde{S}_t$and S_t in this embodiment is based on metrics, more particularly in the equalizers 728a and 728b the respective channel filter taps $\hat{H}$ and $\tilde{H}$ are each used to decode the symbols $\hat{u}$ and $\tilde{u}$. The metrics for the second signal model $\tilde{S}_t$ is similar to equation 24, and the metrics for the first signal model S_t is similar to equation 25, but instead of minimizing with respect to the radio channels $\tilde{H}$ and $\hat{H}$ one minimizes with respect to the symbols $\hat{u}$ and $\tilde{u}$. Thus, the same kind of decision process can be used as in the channel estimation case. For example, if metric $(of\ \tilde{S}_t) < \alpha$ (metric) $(of\ S_t)$ then use the estimated symbols from the second signal model $\tilde{S}_t$ to further process the received signal; otherwise use the estimated symbols from the first signal model S_t to further process the received signal.

It should be understood that the direct conversion receiver 700 can also be adapted to use the internally generated test signals c_t and d_t (see FIGURE 5) or the internally generated test signal g_t (see FIGURE 6) to estimate the parameters $\hat{a}$ and $\hat{b}$.

Referring to FIGURE 8, there is a flowchart illustrating the basic steps of an exemplary method 800 in accordance with the present invention. Beginning at step 802, the first signal model S_t is used to estimate the set of first channel filter taps $\hat{H}$.

At step 804, the second signal model S_r is used to estimate the set of second channel filter taps $\tilde{H}$, including estimating the a and b parameters using an optimizing algorithm such as described above with respect to the first embodiment. The received signal, the internally generated signals c_t and d_t (see FIGURE 5) or the internally generated signal g_t(see FIGURE 6) can be used to estimate the parameters $\hat{a}$ and $\hat{b}$.

At step 806, a selection of the first signal model S_t or the second signal model S_r is made depending on which model better enables the elimination of the near-channel interfering signal from the received signal (see description associated with the first embodiment). The selection of the first signal model S_t or the second signal model S_r can take place before the equalization of the received signal (see FIGURE 4) or after parallel equalizations of the received signal (see FIGURE 7).

In the event the selection of the signal models S_t or S_r takes place before the equalization of the received signal, the selection is made using the estimated plurality of first channel filter taps $\hat{H}$, the estimated plurality of second channel filter taps $\tilde{H}$ and at least one quality parameter. Otherwise, in the event the selection of the signal models S or S_r takes place after parallel equalizations of the received signal, the selection is made using metrics and the decided signals $\hat{u}_t$ and $\tilde{u}_t$.

At step 808, the received signal is decoded and further processed using the selected signal model S_t or S_r.

From the foregoing, it can be readily appreciated by those skilled in the art that the present invention provides an apparatus and method that compensates for the problematic time-varying DC offset by effectively eliminating the AM interferer from the received signal. Also, the apparatus and method disclosed can suppress the AM interferer in a cost and current efficient manner as compared to the prior art. It will also be apparent to workers in the art that the invention can be readily

implemented, for example, by suitable modifications in software, hardware or both, in conventional radio receivers such as direct conversion receivers.

Although several embodiments of the method and apparatus of the present invention have been illustrated in the accompanying Drawings and described in the foregoing Detailed Description, it will be understood that the invention is not limited to the embodiments disclosed, but is capable of numerous rearrangements, modifications and substitutions without departing from the spirit of the invention as set forth and defined by the following claims.

WHAT IS CLAIMED IS:

1. An apparatus capable of compensating for a time-varying DC offset by substantially removing an amplitude modulated interfering signal from a received signal, said apparatus comprising:

a first channel estimator for estimating a plurality of first channel filter taps using a first signal model;

a second channel estimator for estimating a plurality of second channel filter taps using a second signal model; and

a processor for selecting which of the first signal model and the second signal model operates better to substantially remove the amplitude modulated interfering signal from the received signal.

2. The apparatus of Claim 1, wherein said first signal model is represented as:

$$S_t = \tilde{I}_t + j\tilde{Q}_t$$

where $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal.

3. The apparatus of Claim 1, wherein said second signal model is represented as:

$$\tilde{S}_t\left(\frac{\hat{b}}{\hat{a}}\right) = \tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$$

where $\hat{a}$ and $\hat{b}$ are estimated parameters of the amplitude modulated interfering signal, and $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal.

4. The apparatus of Claim 3, wherein said estimated parameters $\hat{a}$ and $\hat{b}$ are determined in accordance with an optimizing algorithm.

5. The apparatus of Claim 4, wherein said optimizing algorithm is executed during each received burst, during a predetermined number of received bursts or during a predetermined number of seconds.

6. The apparatus of Claim 3, wherein said estimated parameters $\hat{a}$ and $\hat{b}$ are estimated using the received signal.

7. The apparatus of Claim 3, wherein said estimated parameters $\hat{a}$ and $\hat{b}$ are estimated using at least one internally generated test signal.

8. The apparatus of Claim 1, wherein said processor operates to select the first signal model or the second signal model depending on which of the signal models has a higher signal-to-noise ratio.

9. The apparatus of Claim 1, wherein said processor operates to select the first signal model or the second signal model using the estimated plurality of first channel filter taps, the estimated plurality of second channel filter taps and at least one quality parameter.

10. The apparatus of Claim 1, further comprising an equalizer for processing the received signal using the selected signal model.

11. The apparatus of Claim 1, further comprising a first equalizer for receiving the estimated plurality of first channel filter taps and a second equalizer for receiving the estimated plurality of second channel filter taps, wherein said processor operates to receive information from said first equalizer and said second equalizer prior to selecting the first signal model or the second signal model.

12. The apparatus of Claim 1, wherein said apparatus is a mobile phone, a base station or a direct conversion receiver.

13. A communications system capable of substantially eliminating a near-channel interferer from a received signal, said communications system comprising:

a first channel estimator for estimating a plurality of first channel filter taps using a first signal model;

a second channel estimator for estimating a plurality of second channel filter taps using a second signal model; and

a processor for selecting which of the first signal model and the second signal model is to be used or was used to substantially eliminate the near-channel interferer from the received signal.

14. The communications system of Claim 13, wherein said first signal model is represented as:

$$S_t = \tilde{I}_t + j\tilde{Q}_t$$

where $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal.

15. The communications system of Claim 13, wherein said second signal model is represented as:

$$\tilde{S}_t\left(\frac{\hat{b}}{\hat{a}}\right) = \tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$$

where $\hat{a}$ and $\hat{b}$ are estimated parameters of the near-channel interferer, and $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal.

16. The communications system of Claim 15, wherein said estimated parameters $\hat{a}$ and $\hat{b}$ are determined in accordance with an optimizing algorithm.

17. The communications system of Claim 16, wherein said optimizing algorithm is executed during each received burst, during a predetermined number of received bursts or during a predetermined number of seconds.

18. The communications system of Claim 15, wherein said estimated parameters $\hat{a}$ and $\hat{b}$ are estimated using the received signal.

19. The communications system of Claim 15, wherein said estimated parameters $\hat{a}$ and $\hat{b}$ are estimated using at least one internally generated test signal.

20. The communications system of Claim 13, wherein said processor operates to select the first signal model or the second signal model depending on which of the signal models has a higher signal-to-noise ratio.

21. The communications system of Claim 13, wherein said processor operates to select the first signal model or the second signal model using the estimated plurality of first channel filter taps, the estimated plurality of second channel filter taps and at least one quality parameter.

22. The communications system of Claim 13, further comprising an equalizer for processing the received signal using the selected signal model.

23. The communications system of Claim 13, further comprising a first equalizer for receiving the estimated plurality of first channel filter taps and a second equalizer for receiving the estimated plurality of second channel filter taps, wherein said processor operates to receive information from said first equalizer and said second equalizer prior to selecting the first signal model or the second signal model.

24. A method of reducing an effect of a near-channel interfering signal on a received signal, comprising the steps of:

utilizing a first signal model and the received signal to estimate a plurality of first channel filter taps;

utilizing a second signal model and the received signal to estimate a plurality of second channel filter taps; and

selecting which of the first signal model and the second signal model operates to substantially remove a near-channel interfering signal from the received signal.

25. The method of Claim 24, wherein said first signal model is represented as:

$$\tilde{S}_t = \tilde{I}_t + j\tilde{Q}_t$$

where $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal.

26. The method of Claim 24, wherein said second signal model is represented as:

$$\tilde{S}_t\left(\frac{\hat{b}}{\hat{a}}\right) = \tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$$

where $\hat{a}$ and $\hat{b}$ are estimated parameters of the near-channel interfering signal, and $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal.

27. The method of Claim 26, wherein said step of utilizing a second signal model further includes estimating said $\hat{a}$ and $\hat{b}$ parameters using an optimizing algorithm.

28. The method of Claim 27, wherein said step of estimating said $\hat{a}$ and $\hat{b}$ parameters occurs during each received burst, during a predetermined number of received bursts or during a predetermined number of seconds.

29. The method of Claim 26, wherein said estimated parameters $\hat{a}$ and $\hat{b}$ are estimated using the received signal or at least one internally generated test signal.

30. The method of Claim 24, wherein said step of selecting the first signal model or the second signal model is determined using the estimated plurality of first channel filter taps, the estimated plurality of second channel filter taps and at least one quality parameter.

31. The method of Claim 24, further comprising the step of decoding the received signal using the selected signal model.

32. The method of Claim 24, further comprising the steps of decoding the received signal using the first signal model and decoding the received signal using the second signal model prior to selecting the first signal model or the second signal model.

33. An apparatus capable of substantially removing an amplitude modulated interfering signal from a received signal, said apparatus comprising:

a channel estimator for estimating a plurality of first channel filter taps using a signal model represented as:

$$\tilde{S}_t\left(\frac{\hat{b}}{\hat{a}}\right) = \tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$$

where $\hat{a}$ and $\hat{b}$ are estimated parameters of the amplitude modulated interfering signal, and $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal; and

an equalizer, coupled to said channel estimator, for processing the received signal using the estimated channel filter taps.

34. A method for substantially removing an amplitude modulated interfering signal from a received signal, said method comprising the steps of:

estimating a plurality of first channel filter taps using a signal model represented as:

$$\tilde{S}_t\left(\frac{\hat{b}}{\hat{a}}\right) = \tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$$

where $\hat{a}$ and $\hat{b}$ are estimated parameters of the amplitude modulated interfering signal, and $\tilde{I}_t$ is a baseband inphase (I) signal and $\tilde{Q}_t$ is a baseband quadrature (Q) signal of the received signal; and

processing the received signal using the estimated channel filter taps.

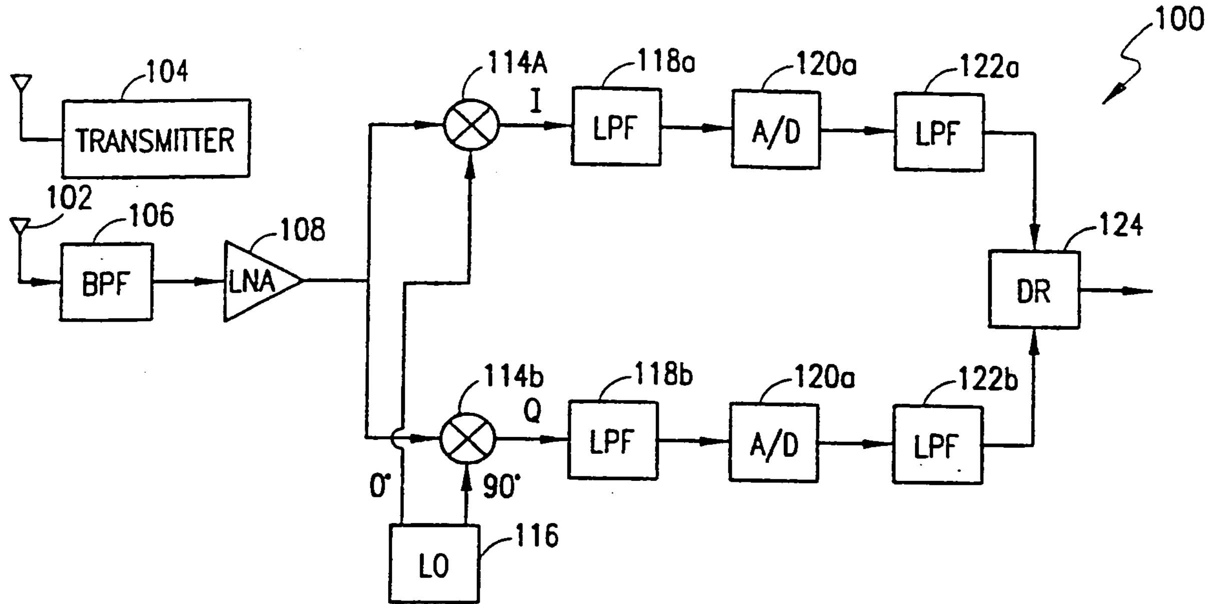

FIG. 1
(PRIOR ART)

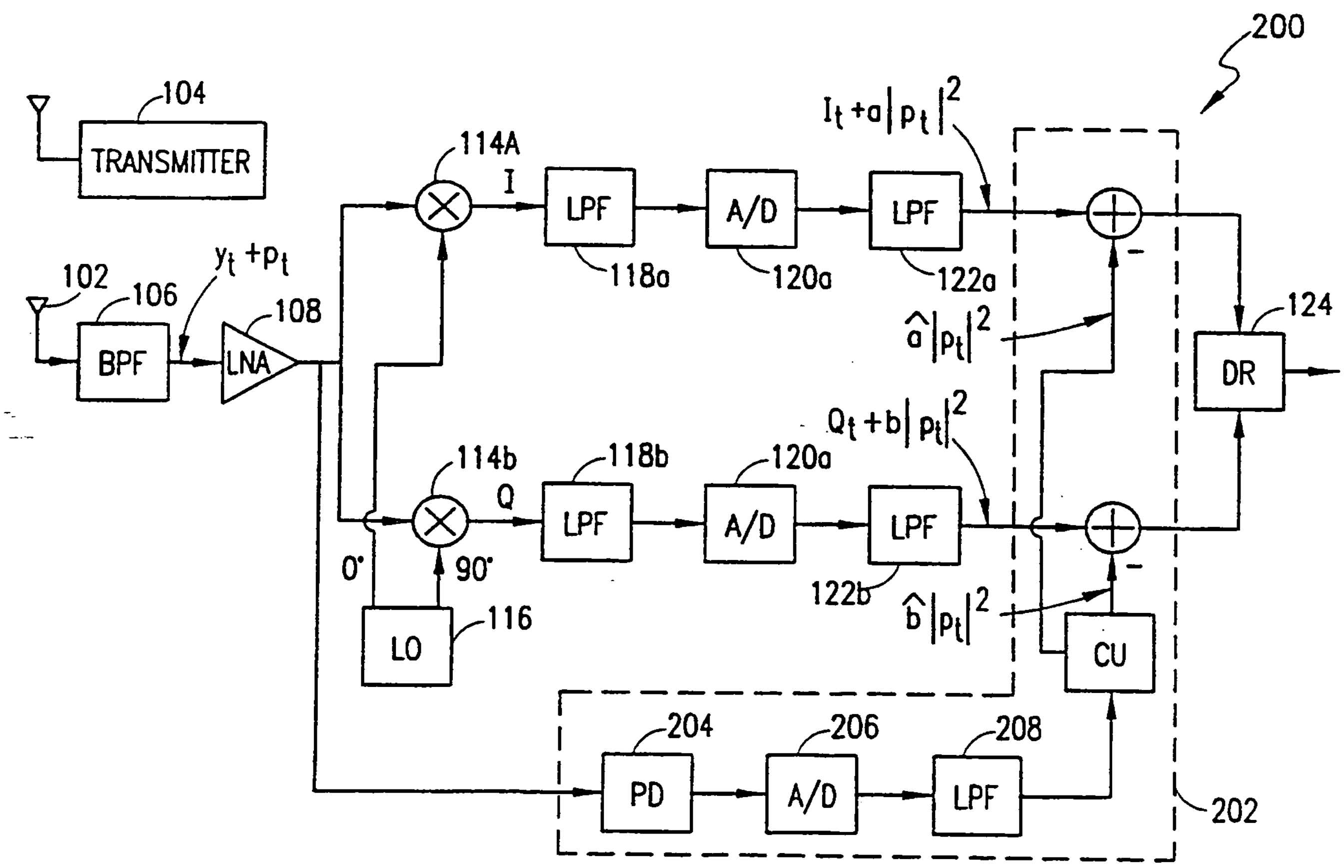

FIG. 2
(PRIOR ART)

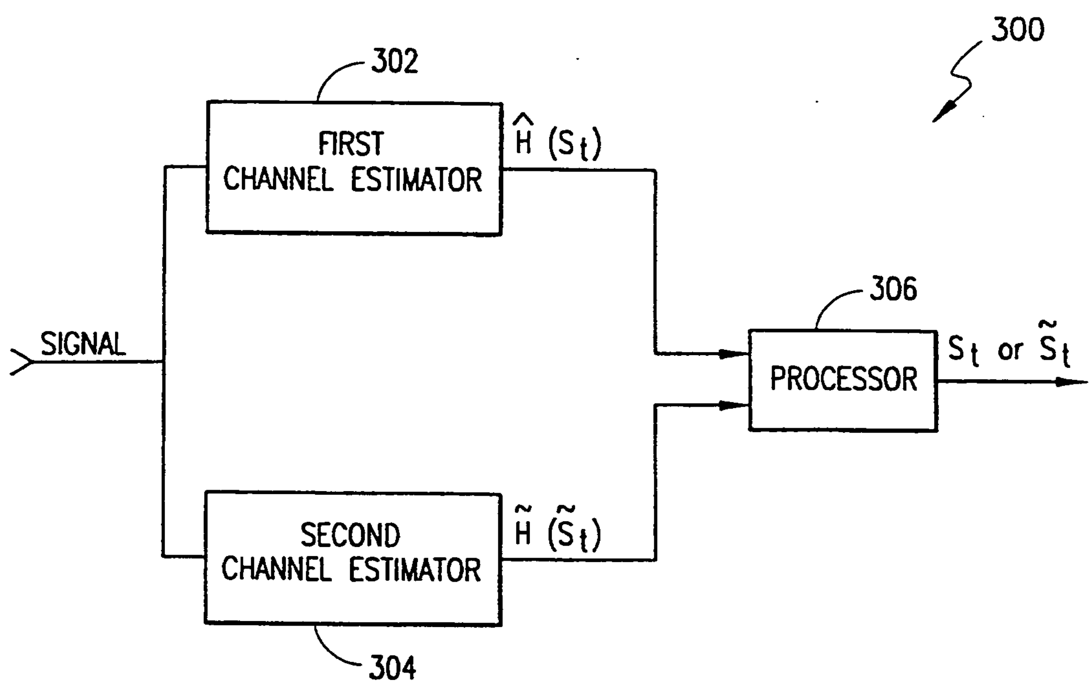
300
302
FIRST
CHANNEL ESTIMATOR
$\hat{H}(S_t)$
306
SIGNAL
PROCESSOR
S_t or $\tilde{S}_t$
SECOND
CHANNEL ESTIMATOR
$\tilde{H}(\tilde{S}_t)$
304

FIG. 3

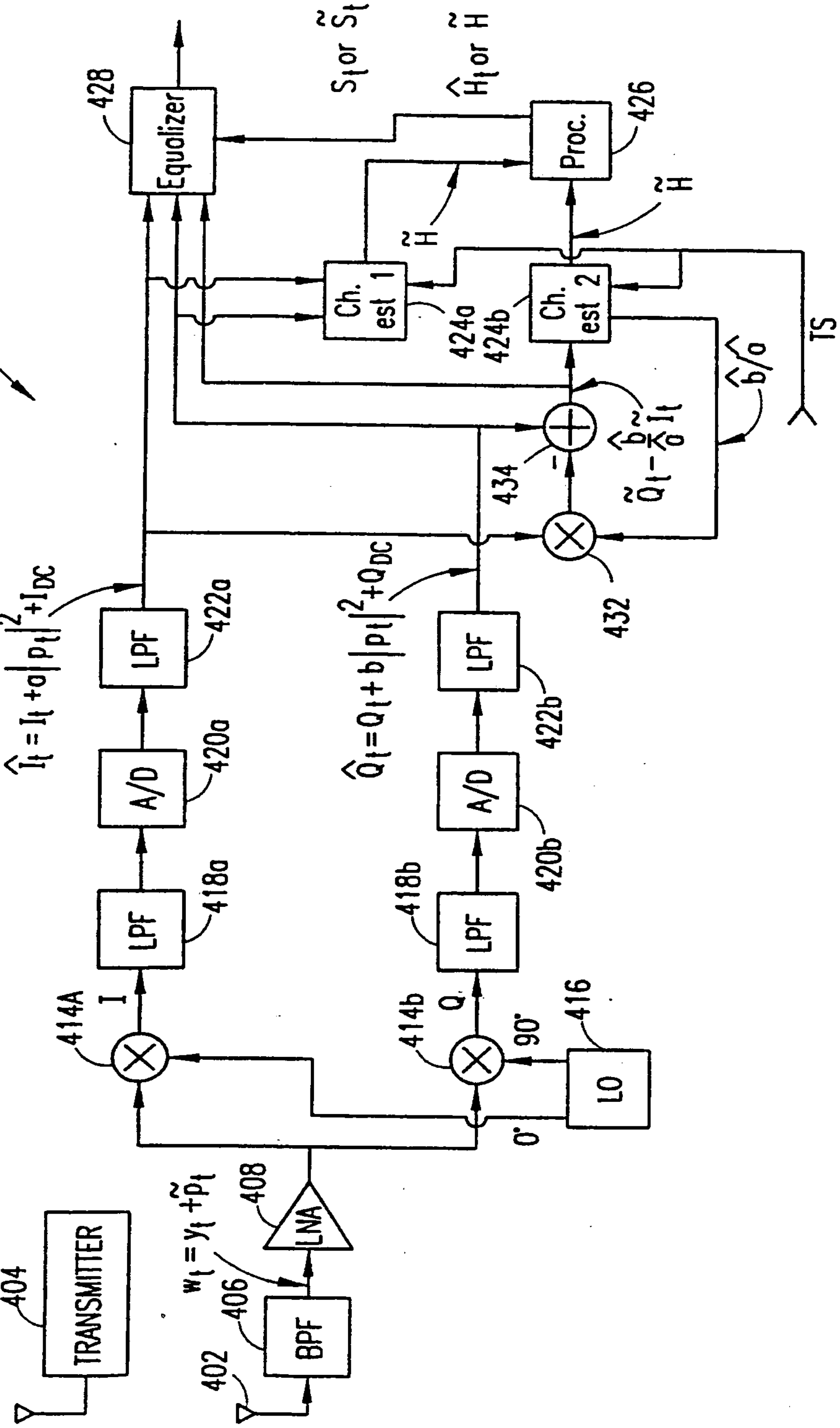
400
404
TRANSMITTER
402
BPF
406
$w_t = y_t + \tilde{p}_t$
408
LNA
414A
I
414b
Q
90°
0°
416
LO
LPF
418a
418b
A/D
420a
420b
LPF
422a
422b
$\hat{I}_t = I_t + a|p_t|^2 + I_{DC}$
$\hat{Q}_t = Q_t + b|p_t|^2 + Q_{DC}$
432
434
$\tilde{Q}_t - \frac{\hat{b}}{\hat{a}} \tilde{I}_t$
$\hat{b}/\hat{a}$
Ch. est 1
424a
Ch. est 2
424b
$\tilde{H}$
Proc.
426
Equalizer
428
$\hat{H}_t$ or $\tilde{H}$
S_t or $\tilde{S}_t$
TS

FIG. 4

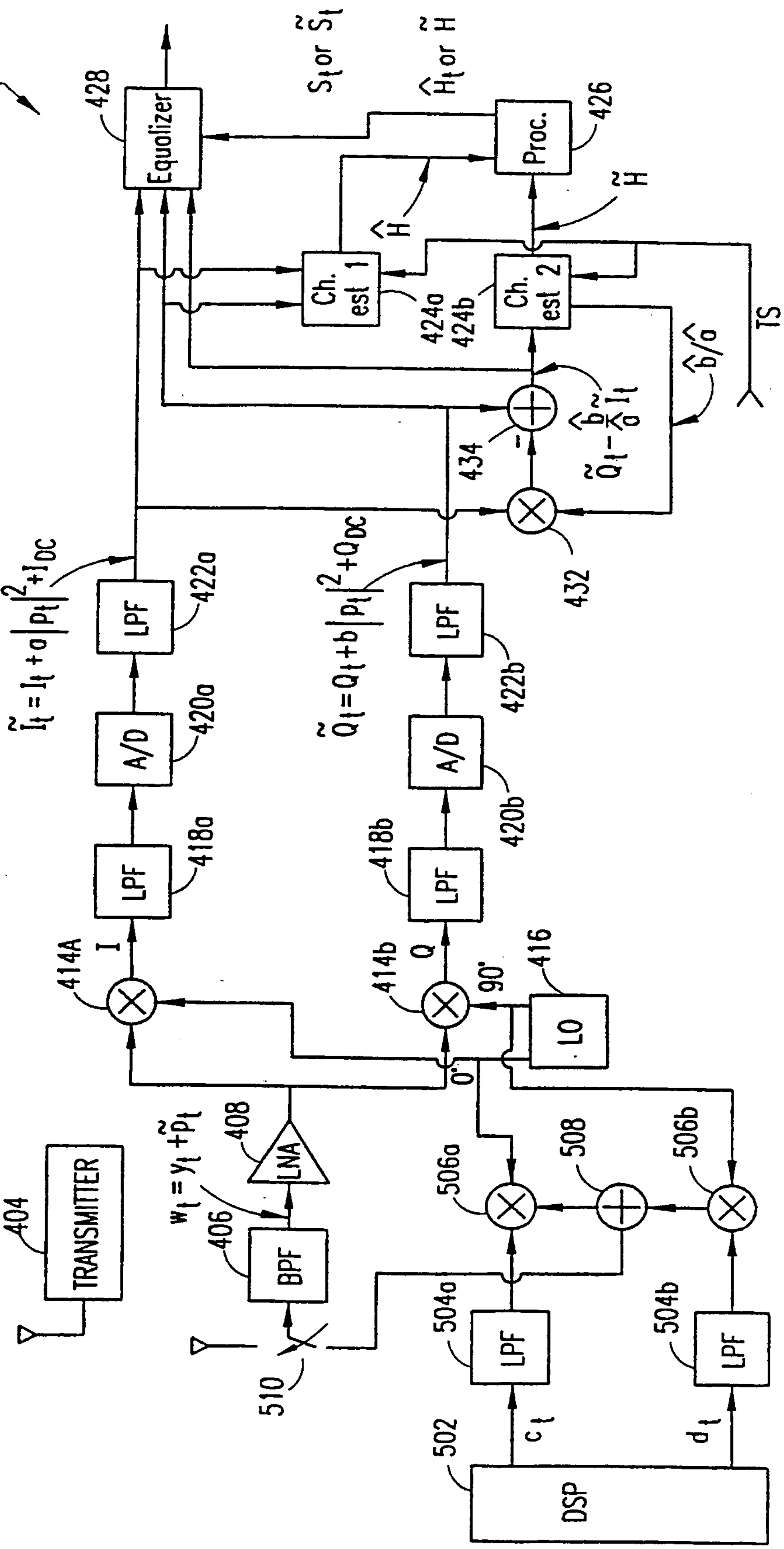
500
TRANSMITTER
404
BPF
406
LNA
408
510
DSP
502
LPF
504a
LPF
504b
506a
506b
508
LO
416
414A
414b
LPF
418a
LPF
418b
A/D
420a
A/D
420b
LPF
422a
LPF
422b
432
434
Ch. est 1
424a
Ch. est 2
424b
Proc.
426
Equalizer
428
TS

FIG. 5

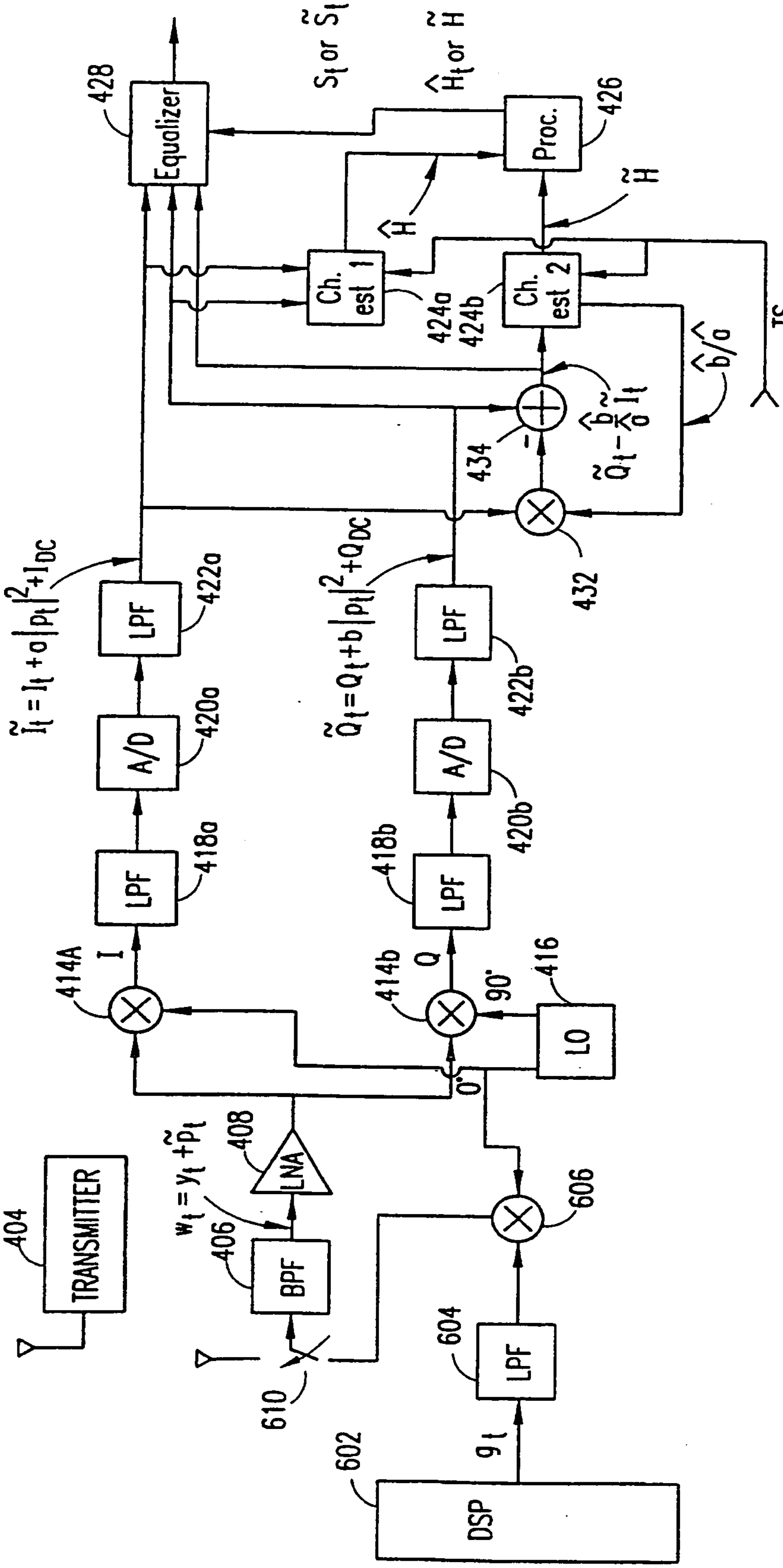
TRANSMITTER
404
BPF
406
LNA
408
610
DSP
602
LPF
604
606
LO
416
0°
90°
414A
414b
I
Q
LPF
418a
418b
A/D
420a
420b
LPF
422a
422b
432
434
Ch. est 1
Ch. est 2
424a
424b
Proc.
426
Equalizer
428
TS

FIG. 6

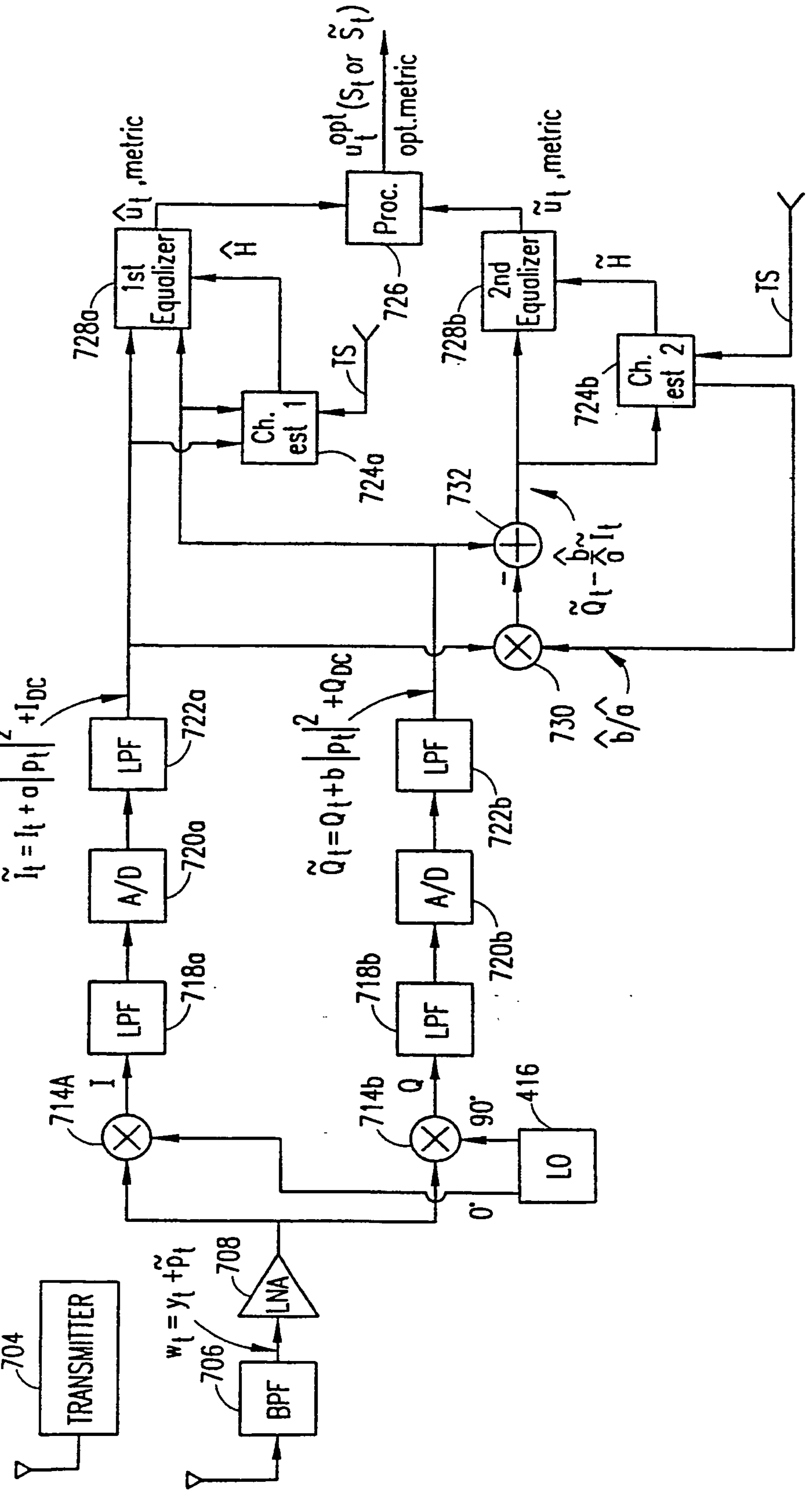
TRANSMITTER
704
BPF
706
LNA
708
$w_t = y_t + \tilde{p}_t$
LO
416
0°
90°
714A
714b
I
Q
LPF
718a
718b
A/D
720a
720b
LPF
722a
722b
$\tilde{I}_t = I_t + a|p_t|^2 + I_{DC}$
$\tilde{Q}_t = Q_t + b|p_t|^2 + Q_{DC}$
730
$\hat{b}/\hat{a}$
732
$\tilde{Q}_t - \frac{\hat{b}}{\hat{a}}\tilde{I}_t$
Ch. est 1
724a
Ch. est 2
724b
TS
TS
1st Equalizer
728a
2nd Equalizer
728b
$\hat{H}$
$\tilde{H}$
$\hat{u}_t$, metric
$\tilde{u}_t$, metric
Proc.
726
$u_t^{opt}(S_t \text{ or } \tilde{S}_t)$
opt.metric

FIG. 7

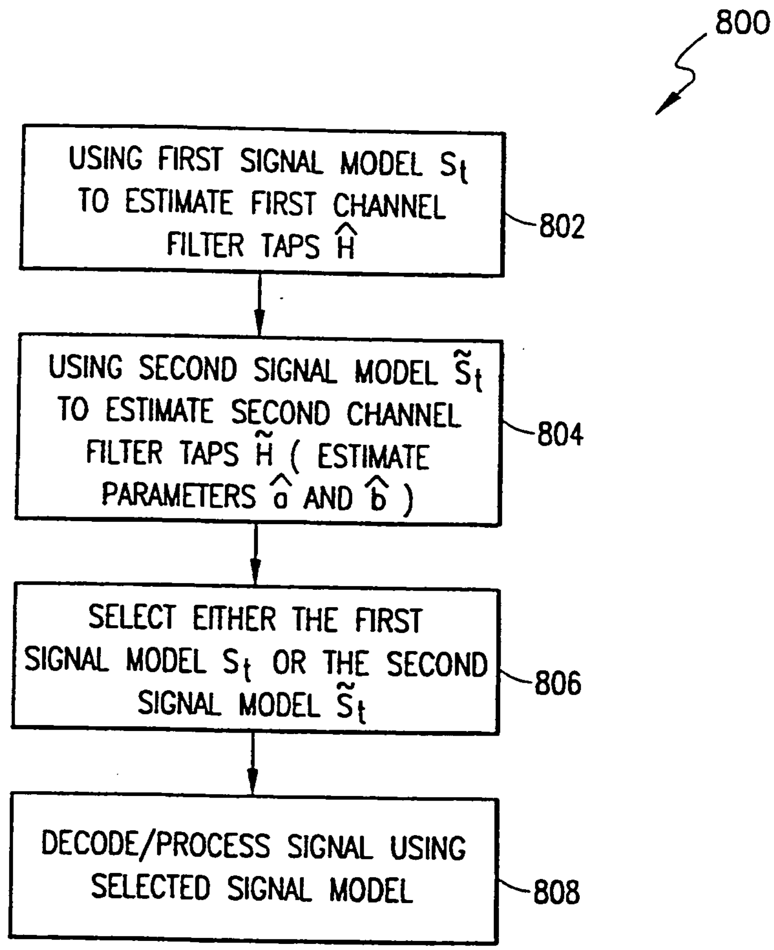
800
USING FIRST SIGNAL MODEL S_t TO ESTIMATE FIRST CHANNEL FILTER TAPS $\hat{H}$
802
USING SECOND SIGNAL MODEL $\tilde{S}_t$ TO ESTIMATE SECOND CHANNEL FILTER TAPS $\tilde{H}$ (ESTIMATE PARAMETERS $\hat{a}$ AND $\hat{b}$)
804
SELECT EITHER THE FIRST SIGNAL MODEL S_t OR THE SECOND SIGNAL MODEL $\tilde{S}_t$
806
DECODE/PROCESS SIGNAL USING SELECTED SIGNAL MODEL
808

FIG. 8

INTERNATIONAL SEARCH REPORT

Intern nal Application No
PCT/EP 00/08715

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H03D1/22 H04L25/06

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H03D H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 5 251 233 A (LABEDZ GERALD P ET AL) 5 October 1993 (1993-10-05) abstract; figure 4 ---	1,9,11, 13,21, 23,24, 30,33,34
A	EP 0 604 208 A (NIPPON ELECTRIC CO) 29 June 1994 (1994-06-29) claims 1,4,5 ---	1,7,10, 13,19, 22,26, 29,31, 33,34

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[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

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"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

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"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
22 December 2000	02/01/2001

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Peeters, M

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INTERNATIONAL SEARCH REPORT

International Application No
PCT/EP 00/08715

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

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A	EP 0 453 213 A (MARCONI GEC LTD) 23 October 1991 (1991-10-23) the whole document	1,13,24, 33,34
A	GB 2 287 620 A (ROKE MANOR RESEARCH) 20 September 1995 (1995-09-20) the whole document	1,13,24, 33,34
A	US 5 422 889 A (SEVENHANS JOANNES M J ET AL) 6 June 1995 (1995-06-06) cited in the application abstract; figure 1	1,13,24, 33,34
A	WO 98 04050 A (ERICSSON GE MOBILE INC) 29 January 1998 (1998-01-29) cited in the application abstract; figure 3	1,13,24, 33,34

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